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Daniel Kürschner

Three phase matrix converter with special control techniques for contactless energy transmission.

6. Power Engineering

Abstract – The use of a matrix converter for contactless energy transmission in the power range of several kilowatts reduces the number of energy conversion steps, avoids electrolytic DC link capacitors, reduces conduction losses in power semiconductors and enables sinusoidal line currents and a four-quadrant operation. In its application as a feeding converter a high output frequency is required (greater than 100 kHz). The transferable electric power and the efficiency of contactless magnetic systems can be considerably improved. Because of the high output frequency, high-grade signal processing elements are needed. The paper investigates special aspects of new control techniques of the three phase matrix converter in combination with contactless energy transmission.

I. INTRODUCTION

Contactless inductive energy transmission technology has been developed within the last few years. Typical applications are assembly robots, machine tools, elevators, linear movable systems and non-contact battery chargers for electric vehicles. By means of the contactless transmission technology, conductor rails, sliding contacts, trailing cables or slip rings can be eliminated. The safety and reliability of the energy supply can be improved. Moreover the limits for speed and acceleration of movable consumers can be increased. Other advantages are: no wear and tear on the electrical contacts, no contact resistance, no spark formation (can be used in explosion-endangered environment) and no non-protected voltage-carrying contacts.

II. MAGNETIC SYSTEM AND MATRIX CONVERTER TOPOLOGY

An inductive power transmission system consists of a primary coil and a static or a movable secondary pick up coil which contains the consumer. Such systems are characterised by a small main inductance and large leakage inductances. The transfer of appreciable electric power requires the compensation of the leakage inductances by resonance capacitors on the primary and secondary coil. With a proper choice of these capacitors, the contactless system represents an ohmic load for the primary inverter. The primary current is nearly sinusoidal and the phase angle between primary voltage and current is zero. Therefore the switching events of the inverter can take place at zero current without auxiliary commutation elements. This zero current switching (ZCS) is an important precondition for reaching higher transmission frequencies. The transmission frequency is the most important electrical parameter for the optimisation of contactless transmission systems. The reachable output power strongly decreases with the air gap length nearly in the same way as

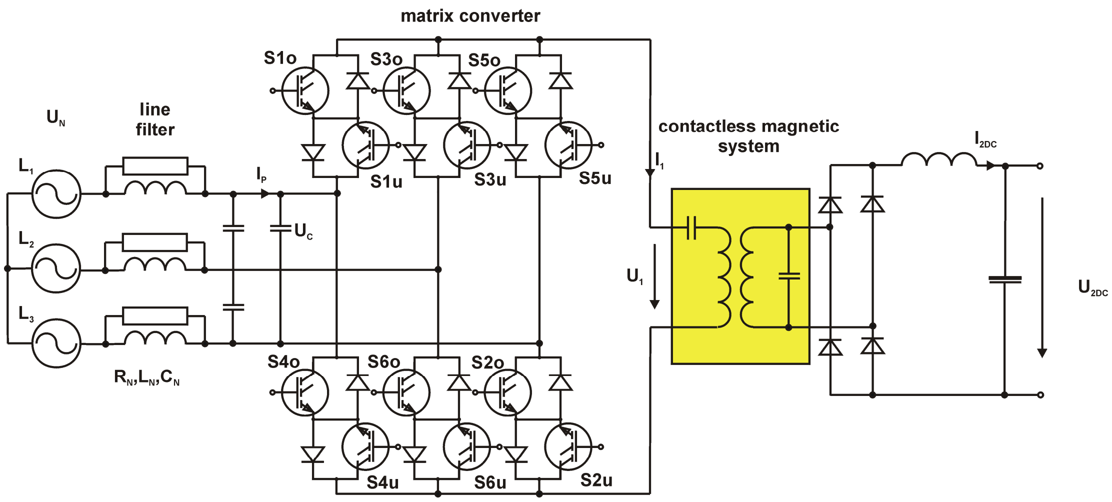

Fig. 1: Matrix converter topology for contactless energy transmission.

the main inductance. On the other hand, by using higher transmission frequencies (greater than 100 kHz), the transferable electric output power and the efficiency of contactless systems can be increased considerably.

By using commonly known multi-step energy conversion concepts (AC to DC, DC to AC), the primary power electronic components produce a great part of the overall losses. Therefore the direct AC to AC energy conversion by means of a matrix converter is an interesting alternative to reduce power electronic conversion losses.

The proposed matrix converter for the power range of several kilowatts is shown in Fig. 1. It is connected to the three-phase network and feeds the primary coil of the magnetic system (f = 100 kHz). The power electronic components on the secondary side produce a constant DC voltage for electrical energy consumers.

III. COMMUTATION STRATEGY

The major difference to conventional feeding converters is the abandoning of a DC-link. Because of the varying amplitude of the feeding lines, they have to change during a line period (see Fig. 2).

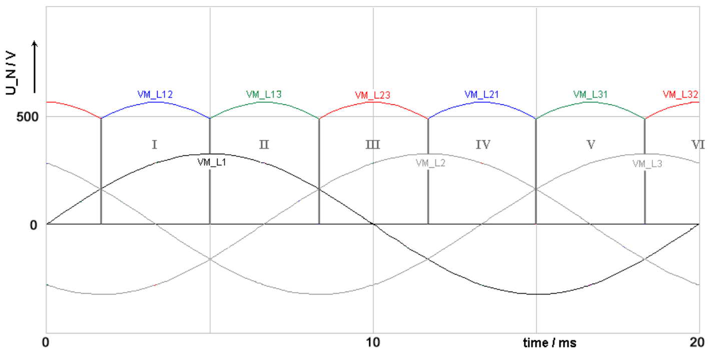

Fig. 2: States of input voltages within one line period.

Within the defined time slots (states I..VI), prearranged lines are used so the three-phase matrix converter operates as a single phase inverter (3 possible circuits). By using e.g. L1 and L2 (time slot I) the IGBTs 1u, 1o, 3o, 3u, 4o, 4u, 6u and 6o are active (Fig. 3).

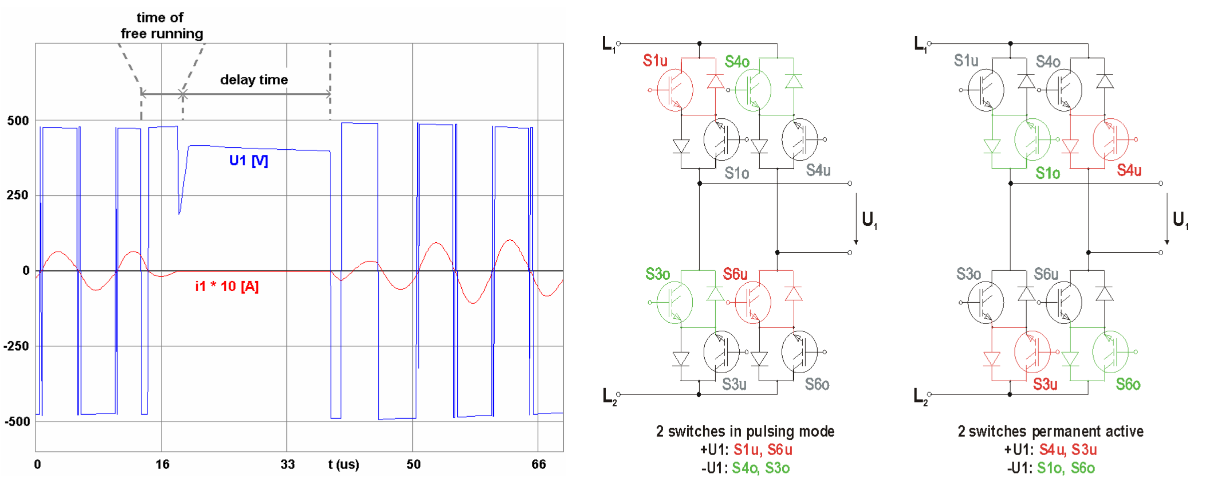

Fig. 3: Commutation during a line switching process (left).; One of three possible rectifier bridges (right).

Two IGBTs switch continuously in alternate mode whereby the adjacent IGBTs are permanently enabled (conductive) so that the current has the possibility to run free (Fig. 3). For the same reason of free running it is necessary to turn off the two types of IGBTs with an offset. After the alternating IGBTs are switched off, the permanently active IGBTs are disabled at a defined time later. Before the next two lines are assigned and the inverting-process continues, a defined delay time should pass to prevent a short circuit.

IV. SINUSOIDAL LINE CURRENTS

A significant problem with power electronic inverters are the harmonics of the line current. To minimise these harmonics, a special control method for the 6 bidirectional power semiconductors (12 anti-parallel IGBT) in combination with optimised line filter parameters can be used. In this way there is no need for an additional power factor correction. To eliminate harmonics in the lower frequency bandwidth, a quasi sine-modulation is used. By electing the feeding lines as shown in Fig. 4 symmetrical line currents over the entire line period can be obtained. For determined lengths and depending on the actual time slot (6 possible states, I..VI) the matrix converter uses either the continuous or the hatched input-lines.

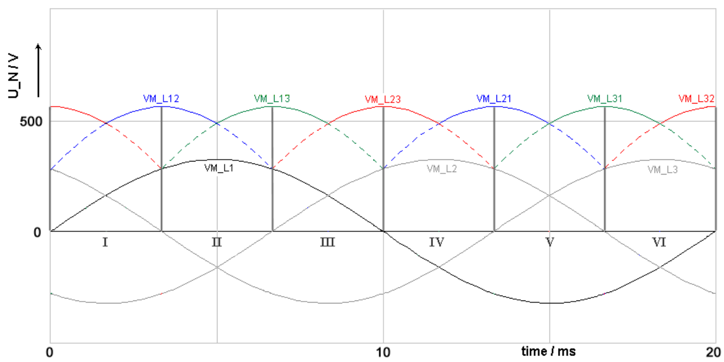

Fig. 4: States of overlapping input voltages within one line period.

The count and the defined order of the pulse lengths result a pulse pattern which allows the following conditions to be kept:

- symmetrical line currents in all phases.
- overall conducting length in a half line period is still 120°.
- the count of current blocks must be uneven.

- the width of the middle block must be equal or greater than 60°.

Because of the used commutation strategy and the delay time, the highest frequency to be damped is related to the count of pulse lengths α within a pulse pattern (20 ms) and so limited by a meaningful count of switching processes within one line period. The harmonics in the upper frequency bandwidth can be damped by small filter elements and ferrite components. Therefore increasing the number of switching processes requires new strategies for line commutation.

The exact length of each block is given by a mathematical equation (1) which is established by the demands of which harmonics $\hat{I}_n$ should be eliminated (*n'th* harmonic).

$$\hat{I}_n = \frac{2 \cdot I_d}{\pi \cdot n} \{2 \cdot \sin n90° \cdot \sin n60° [2 \cdot (\kappa)]\} \tag{1}$$

To eliminate n harmonics (n = 3, 5, 7, 9, 11, …) α_n will be compounded of the following cosine-terms.

$$\kappa_n = (\cos nx_1 - \cos nx_2 + \cos nx_3 - \cos nx_4 + \ldots - \cos nx_n) \tag{2}$$

The terms x_n comprise of sums of specific pulse lengths α_n. With $\hat{I}_n = 0$ you get a system of equations and so the pulse lengths α_1 .. α_n.

To eliminate higher harmonics (*n>11*) the system of equations get very complex. Then there is the possibility to approximately determine values for the pulse lengths by a mathematical comparison of a linear and a triangle function (see Fig. 5).

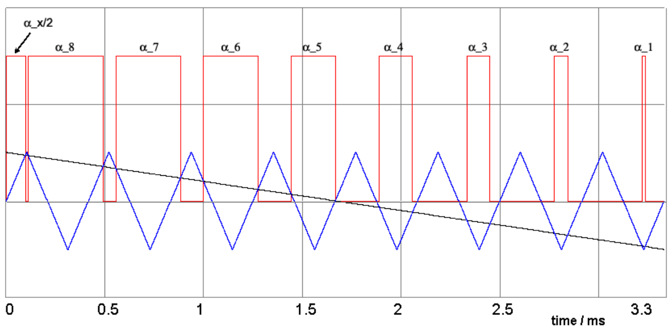

Fig. 5: Alternative determination of the pulse lengths.

Experimental analyses and norm conditions have shown that a block count of 17 blocks and 102 line switching processes per line period obtain good results.

V. SIMULATION

In Fig. 6 (left) the filtered line current at 17-pulse is shown over a half line period. The 100 kHz-ripple is completely extenuated by the filter. To eliminate further and higher harmonics and to improve the sinusoidal form it is necessary to elaborate the parameters of the series dumped line filter shown in Fig. 6 (right). By analysing the admittance functions of the single phase equivalent circuit it is possible to calculate values for L, C an R_P.

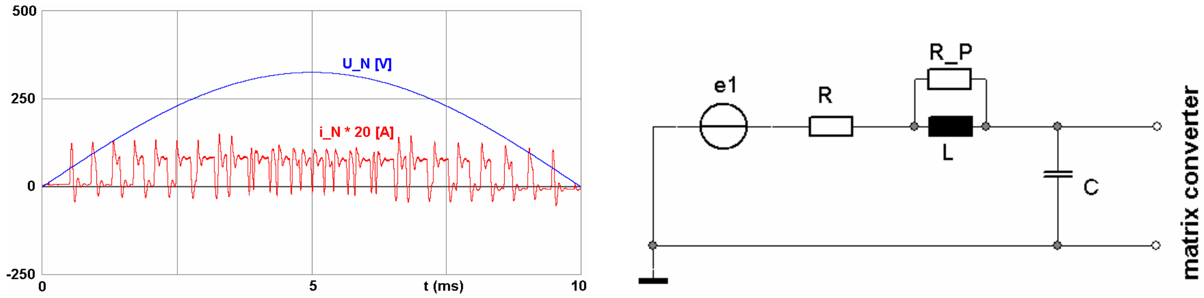

Fig. 6: Filtered line current at 17-pulse modulation (left).; Equivalent circuit of the series dumped line filter (right).

The magnitude spectrum $(|G(s)|)$ of

$$\frac{i_N}{i_P} = \frac{\left(1 + s \cdot \frac{L}{R_P}\right) \cdot \left(1 + s \cdot C \cdot \frac{e}{i_P}\right)}{1 + s \cdot \left(RC + \frac{L}{R_P}\right) + s^2 \cdot LC \cdot \left(1 + \frac{R}{R_P}\right)} \tag{3}$$

at $e = 0$ gives information about the damping effect of the harmonics of i_p by modifying the filter parameters. The influences of parameter variation in view of the capacitive reactive power demonstrates the phase spectrum $(\varphi[G(j\omega)])$ at $e \neq 0$. The value of R depends on the type of line connection and is given by the line-impedance. The adjustment of the resistor R_P should provide an optimum between the value of the filter damping factor and minimising the absorption losses. The resonance frequency and the damping factor have to be chosen in a meaningful range. Here the resonance frequency is given by the minimal occurring frequency caused by the modulated line current. The average current block length at 17-pulse is about 330 µs.

In Fig. 7 (left) the filtered line current at well-tuned passive filter elements and at 17-

pulse is shown over a half line period. The 100 kHz-ripple and the ripple from the sine-pattern is completely extenuated by the filter. The phase difference contrary to the fundamental wave is less than 10°. If C_N is increased, the capacitor voltage is free from further ripples. However the capacitive phase angle would also be enlarged. A compromise between weight, volume of the filter and adherence of norm conditions must be found. Elaborating the parameters of the series dumped line filter (L, C an R_P, Fig. 6., right) occurred by analysing the admittance functions of the single phase equivalent circuit.

Fig. 7 (right) shows the concluding frequency spectrum of the line current at 17-pulse up to 1 kHz. The black bars mean the limits as given in EN 61000-3-2.

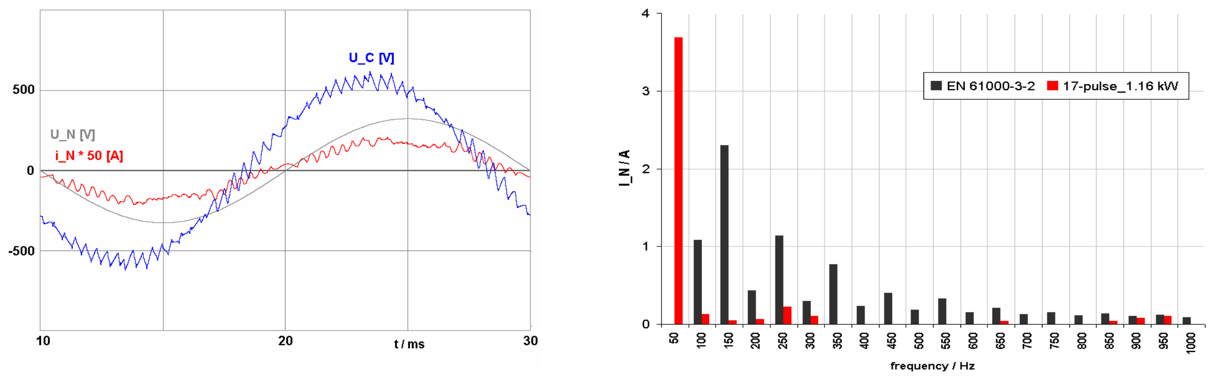

Fig. 7: Line current (I_N) and capacitor voltage (U_C) at well-tuned filter parameters (left); Spectrum of the line current at 17-pulse (right).

VI. IMPLEMENTATION

The requirements of 17 blocks and 102 line switching processes per line period postulate a complex control logic for the 12 IGBTs. Due to high switching frequencies (100 kHz-PWM) and consequently time-critical working conditions, an implementation by a FPGA or a CPLD is reasonable. To implement a real system, the commutation strategy like shown in (III.) was used. The delay time ensures a safe operating condition and offers in addition by increasing or decreasing a regulating variable to control the output voltage U_{2DC}.

To realise a real topology of the three phase matrix converter, fast signal processing components are required and used. An interaction between a 16-bit microcontroller and a high-grade FPGA allows a parallel execution of different tasks. Time critical tasks e.g. the generation of the logic control signals can be processed by the FPGA whereas complex calculations and control operations are suitable jobs for the microcontroller. In this application a Spartan 3 FPGA and an Infineon XC161 Evaluation Board go into action. The user handling is realised via JTAG. More precise, the user controlled input and output variables for the PWM and for the state machine as well as enable and control variables are transferred between the PC and the microcontroller. Computing the pulse lengths for the state machine, AD-conversion, computing regulating variables and synchronisation with the public power supply are further jobs of the microcontroller. To communicate with the FPGA an external data and address bus is also available. The FPGA works as a stand-alone component. It contains the generation of the 100 kHz-PWM signals inclusive a softstart-routine and the proper state machine, which means the processing of the 102 states per line period and the particular choice of the feeding lines. The generation of the 100 kHz-PWM signals necessitates a timing resolution of max. 50 ns so the oscillating frequency of the FPGA must be at least f = 20 MHz. Fig. 8 indicates the transferred variables between the signal processing and the power components.

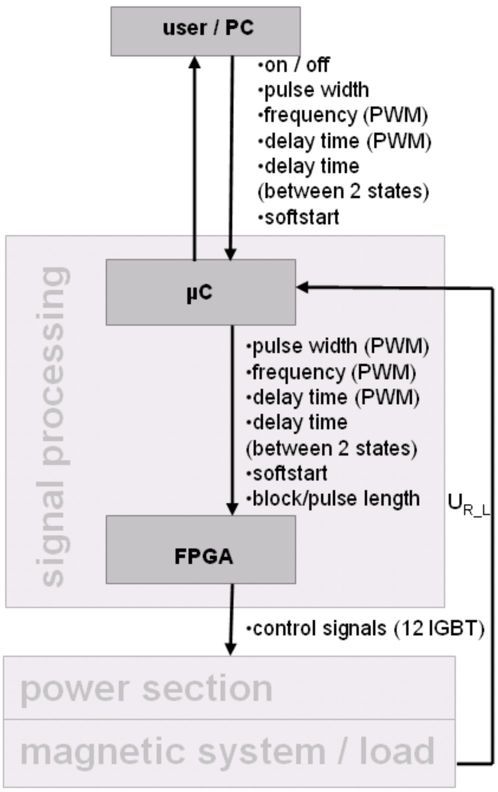

Fig. 8: Transmitted variables between the signal processing and power components.

A block diagram of the entire configuration is shown in Fig. 9 (left) and of the real test platform in Fig.10. The signal processing components as well as the power section is connected via an adjustable transformer because required voltage-proofed IGBTs are not available at this time.

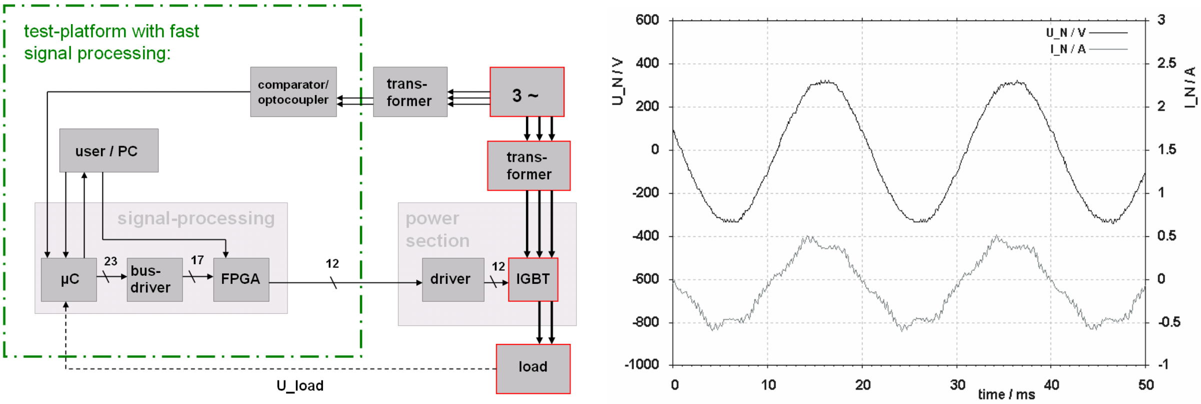

Fig. 10: Block diagram of the entire configuration of the three phase matrix converter (left). Measured line current I_N of the matrix converter (right).

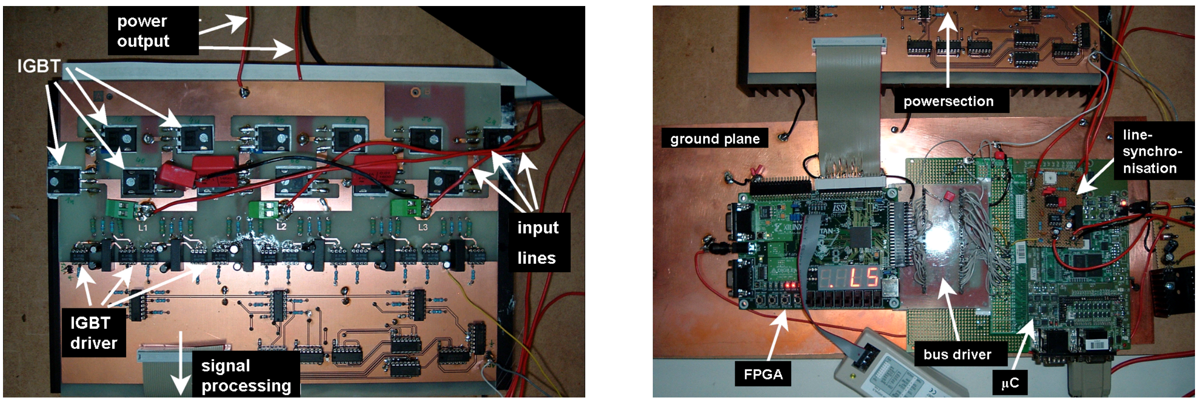

Fig. 11: Power section (left) and signal processing components (right) of the real configuration of the three phase matrix converter.

Measurements at low power offer good results and are conform to the simulation. Fig. 10 (right) shows the line current I_N at line filter parameters as determined in (V.). The capacitive phase angle is nearly zero and the ripple is less than postulated.

VII. CONCLUSIONS

By using higher transmission frequencies (greater than 100 kHz), the transferable electric power and the efficiency of contactless energy transmission systems can be increased considerably. Moreover, high transmission frequencies lead to smaller filter elements and ferrite components. The use of a matrix converter for contactless energy transmission reduces the number of energy conversion steps, avoids voluminous and expensive electrolytic DC-link capacitors, increases the reliability at high temperatures, reduces conduction losses in power semiconductors and enables four-quadrant operation with sinusoidal line currents. The paper investigates special aspects of new selection techniques of the three phase matrix converter in combination with contactless energy transmission. This technique allows sinusoidal line currents so that an additional power factor correction is not necessary. Therefore a new line commutation strategy was found, pulse patterns for the line currents were optimised and a calculation method for the line filter parameters was developed. The simulated behaviour of the three phase matrix converter was verified on the basis of a real configuration. Therefore a combination of a microcontroller and a FPGA was used to generate the logic signals and to determine the pulse pattern. In this way real time signal processing can be performed. In the next step the experimental confirmation will continue at higher input voltages. Further investigations will expand on synthesising a control unit and on related tests of elected regulating techniques with special regulating variables.

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